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A SIMPLE METHOD OF DOPPLERISING A PULSED MAGNETRON RADAR

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KEY TERMS

Doppler radar, magnetron, meteorological radar.

ABSTRACT

We describe a method of Dopplerising an S-band magnetron-based meteorological radar. A digital technique is used to subtract the random phase of the magnetron transmitter from the phase of the return signal. Calibration against a stationary target shows that the phase of the return is constant, with a standard deviation of about 2° and a standard error of <0.1°.

1.INTRODUCTION

An ideal pulsed Doppler radar utilises fully coherent transmission and reception [1]. In a high power Doppler radar, this usually involves using an expensive klystron or TWT amplifier in the transmitter. A less expensive alternative would be to use a magnetron, but a magnetron has a random starting phase and its frequency chirps during the transmit pulse. To overcome this problem many techniques have been developed. One approach is to injection lock the magnetron; in this technique a coherent signal is injected into the magnetron just before it is fired, but the injected signal has to be quite large, typically only 10dB below the magnetron power.

Another technique, known as 'coherent on receive', takes a sample of each transmitted RF pulse and converts it to an intermediate frequency (IF); this IF signal is then used to re-lock the Coherent Oscillator (COHO) to the phase of the IF sample. The phase of the signal reflected from the target is then compared with the phase of the COHO. This leads to a conflict in the specification of the COHO: on the one hand it must be very stable if the phase of the return signal is to be recovered accurately, but on the other hand it must be possible to change the phase of the oscillator to re-lock it during the very short (typically < 1 μsec) transmitted pulse. In addition, it is not obvious which part of the transmitted pulse the

An analog technique [2] has been used to correct for the random phase of the magnetron, but a digital approach has several advantages. Firstly, in a digital system, one can easily remove the DC offset in the phase detectors and also balance the amplitudes of the I Q channels and so reduce the error in angle measurements. Secondly, digital processing of the I Q data simplifies the circuitry and the mathematical computation will be more accurate than for an analogue technique. Finally, this digital technique does not involve any modification of the existing radar RF system; it requires only a sample from the IF signal of the receiver and the addition of a mixer to obtain the phase of transmitted microwave pulse.

In Section 2 we summarise the principle of the new method, and in Section 3 we describe the hardware specifications for the various elements such as the COHO, the STALO, amplifiers and phase detection circuitry. In Section 4 we present measurements of the phase characteristics of the magnetron transmitter, and Section 5 discusses results of radar dwells on ground clutter in which the phases reflected by a stationary target have a standard deviation of about 2° and the one second dwell time yields a standard error of 0.1°. Spectral analysis shows that the zero frequency spectral componenent is 50dB above the system noise floor.

In Section 5 we present some measurements of the specific differential propagation phase shift, K_{DP}, a new parameter which can be used to characterise precipitation. When the radar wave travels through a region of horizontally aligned oblate precipitation particles, the phase of the horizontally polarised wave progressively lags behind the vertically polarised one. The new technique allows this phase difference to be measured more accurately than previously reported.

2. THE PRINCIPLE OF ECHO PHASE ACQUISITION

Figure 1 summarises the principle of operation of the magnetron based Doppler radar. We assume a transmitted signal of carrier frequency f_o, and a pulse modulation function A_{Tx}:

$$T_x = A_{Tx}\exp[-j2\pi f_O t + j\varphi_i] \qquad (1)$$

where φ_i is the random phase angle for the ith pulse and t is the elapsed time from the start of each transmission. The return signal is then

$$R_x = A_{Rx}\exp[-j2\pi f_O(t-2r/c) + j(\varphi_i+\zeta_i)] \qquad (2)$$

where r is the range of the target, c is the velocity of light, and ζ_i is phase change on scattering from the target for the ith pulse.

Mixing these two signals, T_X and R_X, with a Stable Local Oscillator (STALO) of frequency f_S and relative phase θ_i, gives two intermediate frequency (IF) signals, which, after low-pass filtering to select the difference frequency are given by::

$$IF_{Tx} = A_{Tx}\exp j[2\pi(f_O - f_S)t + (\varphi_i - \theta_i)] \tag{3}$$

$$IF_{Rx} = A_{Rx}\exp j[2\pi(f_O - f_S)t - 4\pi f_O r/c + (\varphi_i + \zeta_i - \theta_i)] \tag{4}$$

The two intermediate frequency outputs are then sent to the phase detectors and low pass filters which compare their phase with the coherent oscillator (COHO), frequency f_C and relative phase ξ_i, to give two signals:

$$\Phi_{Tx} = 2\pi\,(f_O - f_S - f_C)t + (\varphi_i - \theta_i - \xi_i) \tag{5}$$

$$\Phi_{Rx} = 2\pi(f_O - f_S - f_C)t - 4\pi f_O r/c + (\varphi_i + \zeta_i - \theta_i - \xi_i) \tag{6}$$

Finally, if the transmitted signal is sampled at time t_1, and the received signal at time t_2, then the 'corrected' phase of received signals is given by

$$\Phi_i = \Phi_{Rx} - \Phi_{Tx} = 2\pi\Delta f(t_2 - t_1) - 4\pi f_O r/c - \zeta_i \quad , \tag{7}$$

$$\textit{where,} \qquad \Delta f = f_O - f_S - f_C$$

In the Doppler processing we require the phase change between successive pulses at a specific range gate, and the subtraction removes the constant Δf term; a similar argument applies for the differential phase shift which also involves the subtraction of phases between successive pulses. Ideally, ΔF, the difference of the COHO and the IF signals should be zero, but this is not necessary.

3. PRACTICAL DETAILS OF THE DOPPLER RADAR

The hardware described is implemented on the Chilbolton radar, operated by the Radio Communication Research Unit of the Rutherford Appleton Laboratory. The magnetron used in these studies is the M509A, manufactured by EEV operating at 3.0765 GHz as

incorporated in the AR-1 radar manufactured by Plessey with a modulator to give a 0.5 μsec long pulse (75m range resolution) and a peak power of 500kW at a pulse repetition frequency of 610Hz. A 3.0465GHz STALO gives an IF of 30MHz. Pulses are transmitted with alternate horizontal and vertical polarisation. A phase change of 180° between successive pulses of the same polarisation is equivalent to a target velocity of 7.5m/s and a Doppler shift of 150Hz. A change of 1° corresponds to a velocity of 0.04m/s. When designing the system, the goal was to measure the return phase to an accuracy of 1°.

The transmitted pulse was sampled with a pyramidal horn, which was placed next to the feed horn in a cabin at the dish focus, with its side at 45° to the horizontal, so that it could serve as a transmit pulse sampler for both polarizations. The output of the horn antenna was firstly attenuated 20 dB and then mixed with the STALO. The output of the mixer (IF_{Tx} - Equation 3) and the IF signal from the receiver (IF_{Rx} - Equation 4) were sent by two long (>50m) coaxial cables to the control room where the dual phase detector board, the COHO and the timing circuits for the sampling were located.

The COHO consists of three parts. The VDS-256 is an inexpensive synthesized clock generator (square wave), using an on-board 10 MHz reference and PLL technique, it has a frequency range of 1 MHz to 255.999 MHz with a frequency resolution of 1 KHz. The output is a 300 mV peak to peak square wave into 50 Ω. Two series low-pass filters are used to eliminate the harmonics of the desired fundamental. After the filtering, very pure sine waves are obtained. Spectral analysis shows that the harmonics are reduced to about 50 dB below the fundamental. Subsequent amplification increases the power to around 15 dBm, the optimum drive level for the phase detector module. The performance of the Doppler radar is unchanged if an HP 8660C frequency synthesizer is substituted for this COHO, confirming satisfactory performance.

The circuitry for detecting the phase of the Tx and Rx signals is shown in Figure 2. The design of this board is critical if both channels are to measure relative phases to an accuracy of 1°. For each I Q channel circuit, symmetry is important so that the same electrical path length is maintained and the two channels are phase-balanced. To reduce mismatch, all connections between devices are by 50Ω microstrip transmission lines. The filters at the IF inputs of the two channels are used for eliminating the IF signal distortions which are mainly from harmonics. These filters are low-pass, having a 3 dB pass-band of 32 MHz, and 40 dB attenuation at 56 MHz. Consequently, they are able to suppress the IF harmonics to better than 40 dB below the desired 30 MHz signal. The low pass filters at the output of the phase detectors are used both to reject the high frequency mixing products of the phase detectors and to limit the bandwidth of the output noise. These output filters have a 3 dB passband of 10.7 MHz and provide 40 dB attenuation at 19 MHz. The bandwidths of all filters are chosen

to be sufficiently broad so that 75m range gates can be resolved without introducing any phase distortion.

Power splitters were chosen to have a phase imbalance between output ports of less than 1° together with good amplitude balance. Phase measurements using a Hewlett Packard Vector Voltmeter HP 8508A confirmed that the IQ technique described above was able to measure the relative phase of either input to better than 1°. Absolute phase measurements are unimportant. As shown in Equation 7, we require only the difference in the phase of the return at each range gate from one pulse to the next. The outputs of the phase detectors are only $\pm$100mV peak to peak, so subsequent amplification (34 dB with a bandwidth of 6 MHz and an input impedance of 50 Ω) was used to provide an appropriate output for the analog to digital converters.

The requirement to estimate phases to 1° affects the specification of other components. The ADC used for sampling Φ_{Tx} has an aperture jitter of 0.5nsec, so the frequency of Φ_{Tx} can be as high as 2 MHz but can still be estimated to considerably better than 1°. The IF bandwidth places an upper limit of 2MHz on the value of ΔF. The aperture jitter of the ADC sampling Φ_{Rx} (Equation 6) is only 60ps, so the sampling gate moves in space by less than 1cm (one part in 10,000), thus ensuring that the correlation of the two co-polar returns can be measured to high accuracy. The sample time t_1 was adjusted to be in the middle of the transmit pulse. For a maximum range of 150km, the elapsed time, t_2, can be as large as 1msec. In this time the phase noise of the STALO should be less than 1°. This leads to a frequency stability requirement for the STALO of 1 part in 10^9, or 3Hz. The magnetron frequency chirp for successive pulses also needs to be reproducible to 3Hz. The figures for the COHO are less stringent, being only 1 part in 10^7.

4. THE PHASE CHARACTERISTICS OF A MAGNETRON PULSE

A first estimate of the phase variation within the transmitted pulse can be obtained from either the I or Q output of the Tx phase detector. Figure 3 displays this phase variation for four different transmitted pulses. Because the IF is amplitude limited, a complete cycle in a trace represents a phase change of 360°. It can be seen that rapid frequency changes occur during the starting and decaying parts of the pulse, but during the high power period, the changes are relatively small, totalling about 30°. Such a 30° phase change during a pulse length of 0.5 μsec is equivalent to a total chirp of about 150KHz. The different traces in the Figure reflect the random starting phase, but two pulses have been chosen with approximately the same starting phase. The accuracy of the phase measurement system is limited to about 1° by factors such as the phase noise in the local oscillators, the frequency dependent scale factor of the phase detectors and the noise figure of the testing system. It is clear that,

within the measurement accuracy, the pulses with the same starting phase have an identical phase variation, implying that the chirp is very regular; a change of 1° indicating that the chirp is reproducible to better than 5.5KHz. This is the smallest frequency change measurable by this technique, and represents an upper limit to the estimate of reproducibility of the frequency chirp. In the next section we will see that the actual performance is much better.

5. PERFORMANCE WITH GROUND CLUTTER AND PRECIPITATION TARGETS

The most direct method to confirm the phase stability of the magnetron Doppler system is to use stationary targets, as shown in Figure 4 for a target at 21.3km range. The data were collected for an antenna dwell of 1.6 seconds, that is 1000 pulses in all, 500 for each polarisation. The return phases are constant with standard deviations for each polarisation of about 2°, corresponding to a velocity uncertainty of about 0.1m/s. The standard error of the mean for the 500 pulses is only 0.1°, equivalent to a velocity of less than 5mm/s. A spectral analysis of the phases is plotted in Figure 5 and demonstrates that the power density around zero frequency is 40dB higher than that at 1.2Hz and 50dB higher than the other components.

To obtain the clutter data in Figures 4 and 5, it is important to choose a day when atmopsheric conditions are very calm. The 2° standard deviation for a target at around 20km implies that the magnetron chirp from pulse to pulse is reproducible to better than 50Hz. Again, this is an upper limit because the observed phase changes could be caused by phase noise in the COHO or STALO, or by variable atmospheric propagation and target movement.

Figure 6 displays a typical phase progression and power spectrum for a meteorological target. The mean Doppler frequency is near 50Hz (about 2.5m/s) with a Doppler width of about $\pm$10Hz (0.5m/s). It is our experience that the Doppler width of meteorological targets rarely falls below 0.5m/s and that, as a result, we conclude that any imperfections in the magnetron phase measurement system are far less than those caused by the intrinsic Doppler width of meteorological targets.

The data in Figure 4 show that the mean phase measured with the vertical and horizontal polarisations differ by about 4°. This difference, called Φ_{DP}, is caused by differing electrical path lengths between the antenna and target for the two polarisations. As the radar wave propagates through precipitation particles which are oblate and aligned with their major axes horizontal, then the horizontally polarised wave becomes progressively delayed and Φ_{DP} increases with range. Figure 7 shows measurements of Φ_{DP} recorded in light rain, with the

radar reflectivity, doppler velocity and Doppler width. The data for a 0.4 second dwell (128 pulses of each polarisation) along a 37.5km path recorded at 512 gates each 75m wide. The radar reflectivity factor (Z) is about 20dBZ and quite uniform; which is equivalent to a rainfall rate (R) of about 0.68mm/h obtained from an empirical relationship [3] of the form $Z = 200R^{1.6}$. In this light rain the values of Φ_{DP} increase very slowly with range. If we average the value of Φ_{DP} over 32 adjacent gates, that is 2.4km range, then the average value at range 19km is -0.43° with a standard deviation of 0.83° and a standard error of 0.15°. At 53km the average value of Φ_{DP} is 0.40° with a standard deviation of 0.75° and standard error of 0.13°. The average gradient of Φ_{DP} is called K_{DP}, which over this 32km range is 0.026 ± 0.006 deg/km. For rain with a Marshall-Palmer drop size distribution [3], the rainrate R is almost linearly related to K_{DP}. The theoretical relationship [4], K_{DP} (deg/km) $= 0.03\ R^{1.15}$ (mm/hr), yields an average rainfall rate of 0.87 ± 0.25 mm/hr over the 32km range, which is consistent with the estimate made from the reflectivity. Further discussion of rainrate estimates from phase measurements is provided in [5].

5. SUMMARY AND CONCLUSIONS

The method introduced in this paper has two advantages: very low cost and high accuracy of phase measurements. The dopplerisation of a magnetron transmitter is achieved without changing the microwave hardware. The standard deviation of phase measured by the system from a stationary target at 21km is about 2° with a standard error of only 0.1°. For the Chilbolton radar, this standard error is equivalent to a Doppler velocity accuracy of 5mm/sec. For meteorological targets, the differential phase shift in rain can be measured to better than 0.001 deg/km, and provide rainfall rate estimates down to 1mm/hr. The residual phase noise of this magnetron system is less than the Doppler width of most meteorological targets, so the use of a klystron would not lead to any improvement in the performance.

ACKNOWLEDGEMENT

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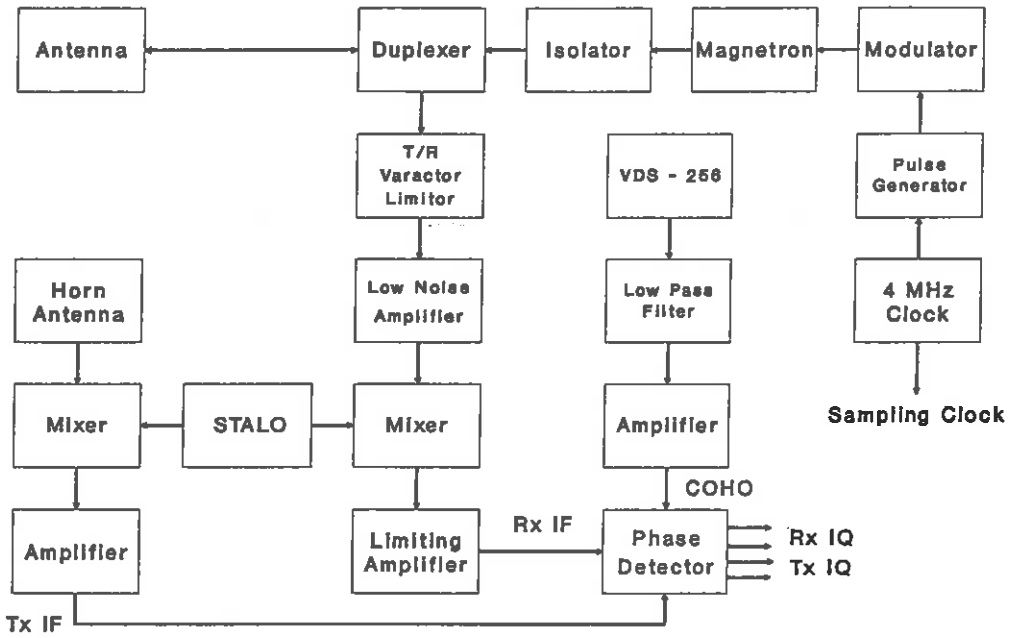


Figure 1 Operating principle of the magnetron Doppler radar

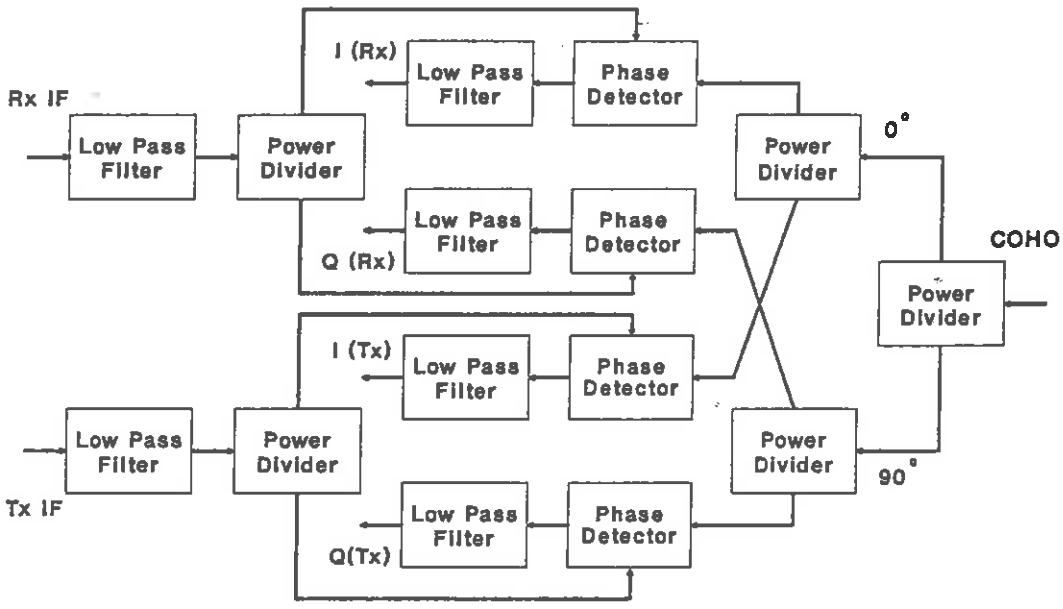
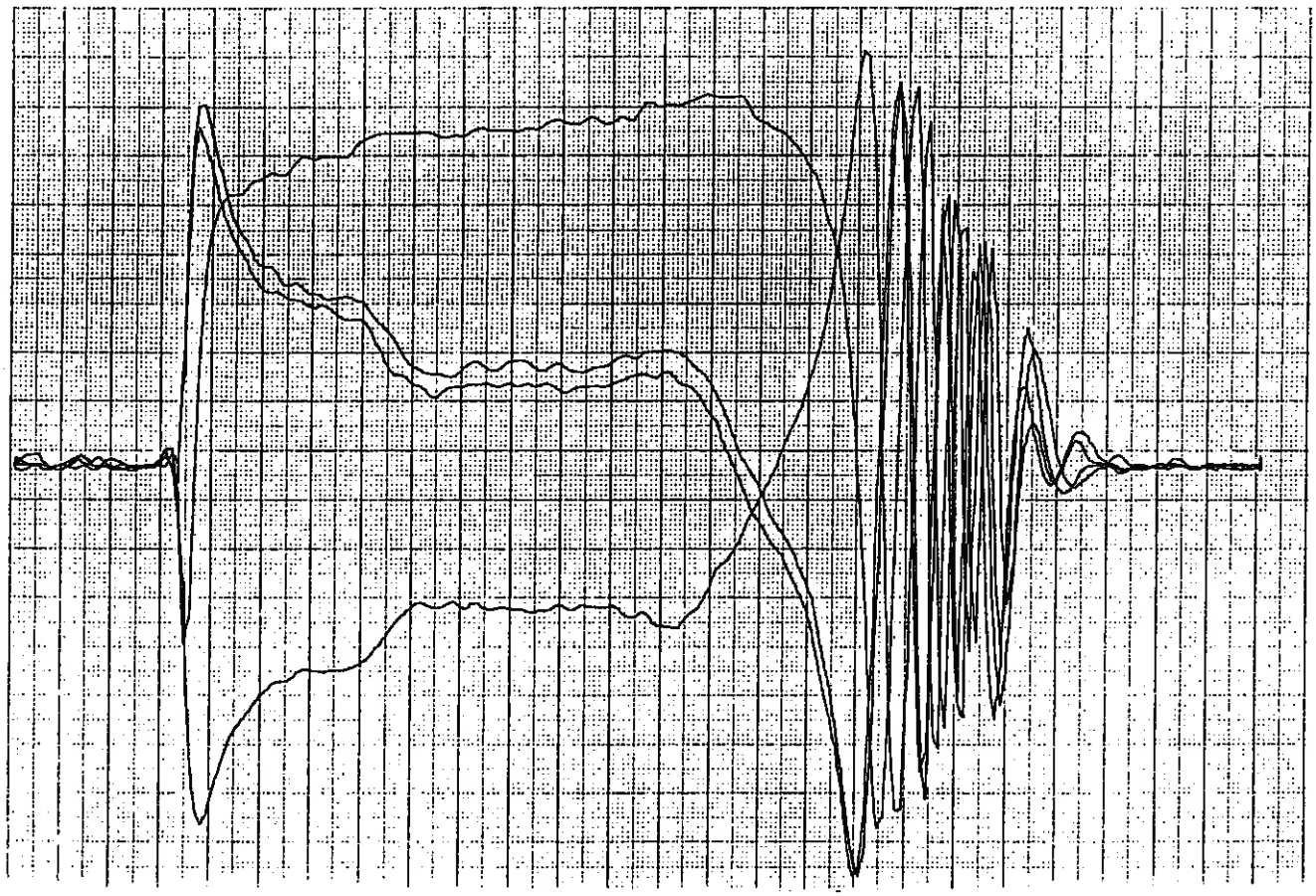


Figure 2 The dual I Q phase detector board

Figure 3 Phase variation during four different pulses transmitted by the magnetron. One cycle represents a phase change of 360°. The phase change during the high power transmission is only 30°.



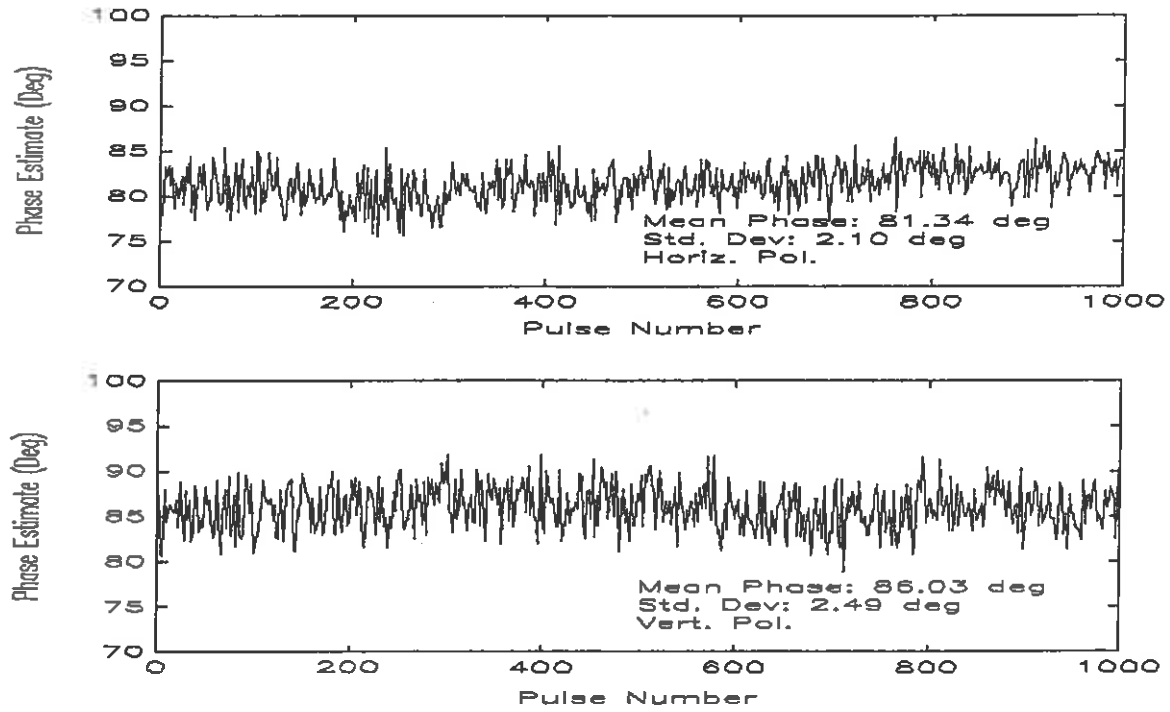


Figure 4 The phases of the echo from a stationary target 21.3km range.

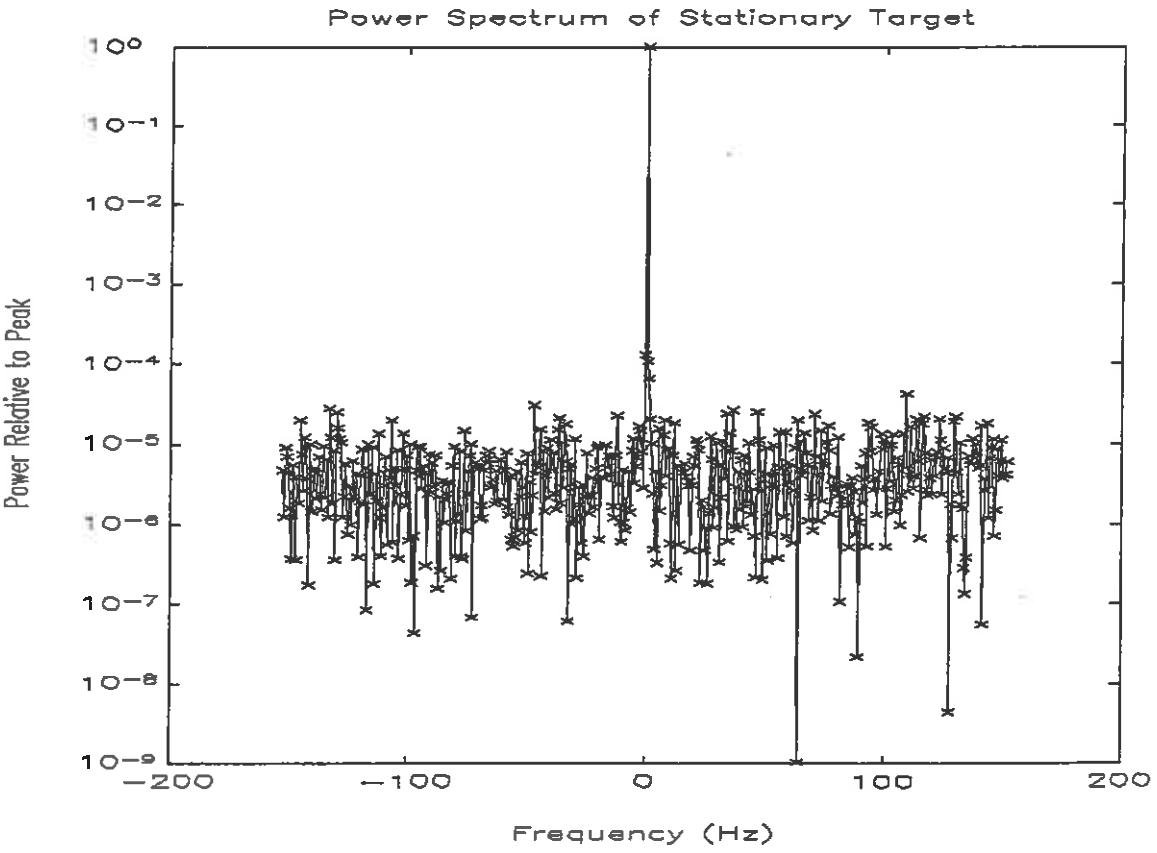


Figure 5 The power spectrum of the stationary target phases in Figure 4.

Figure 6 Phase progression and power spectrum for the return at a single gate from a meteorological target.

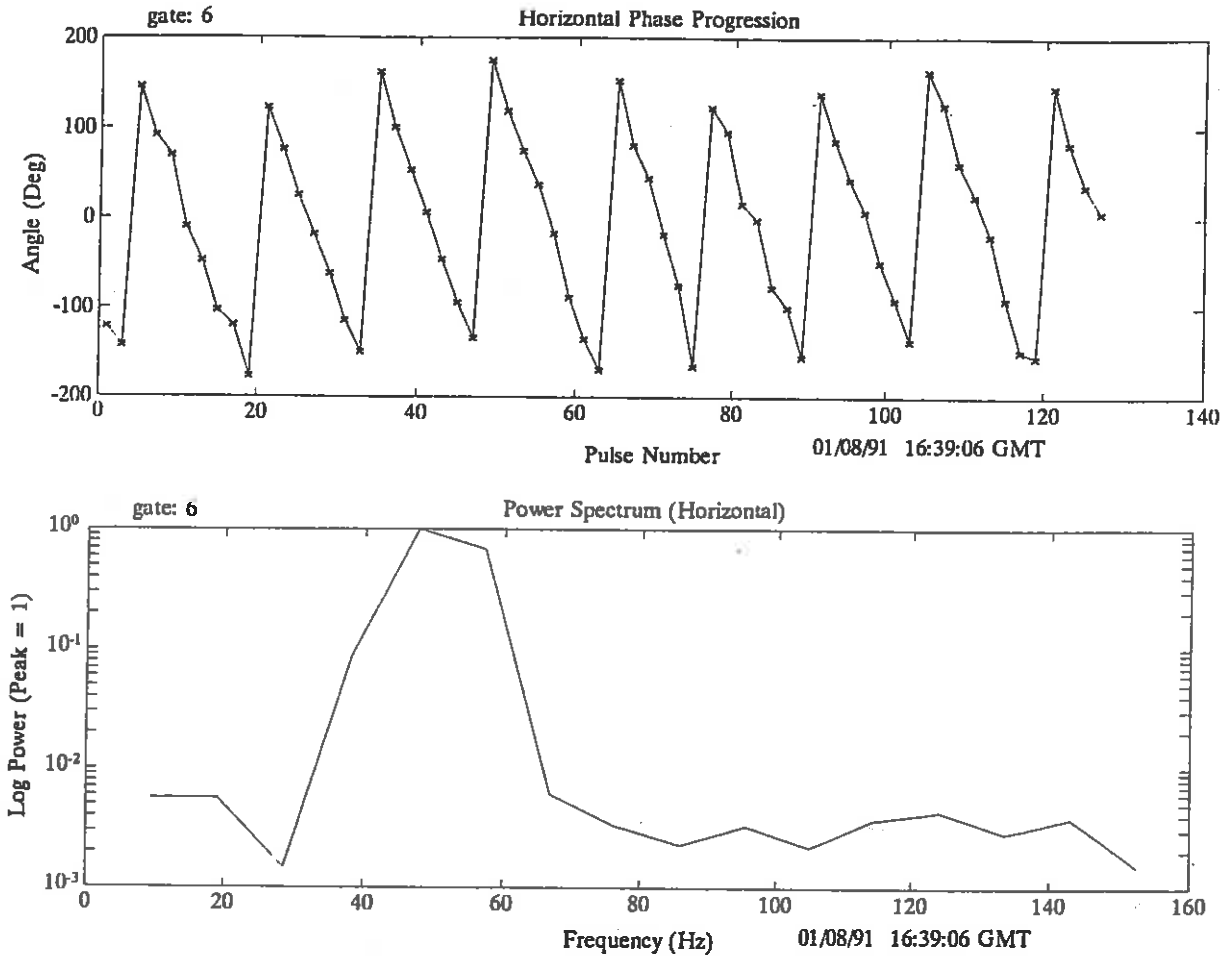
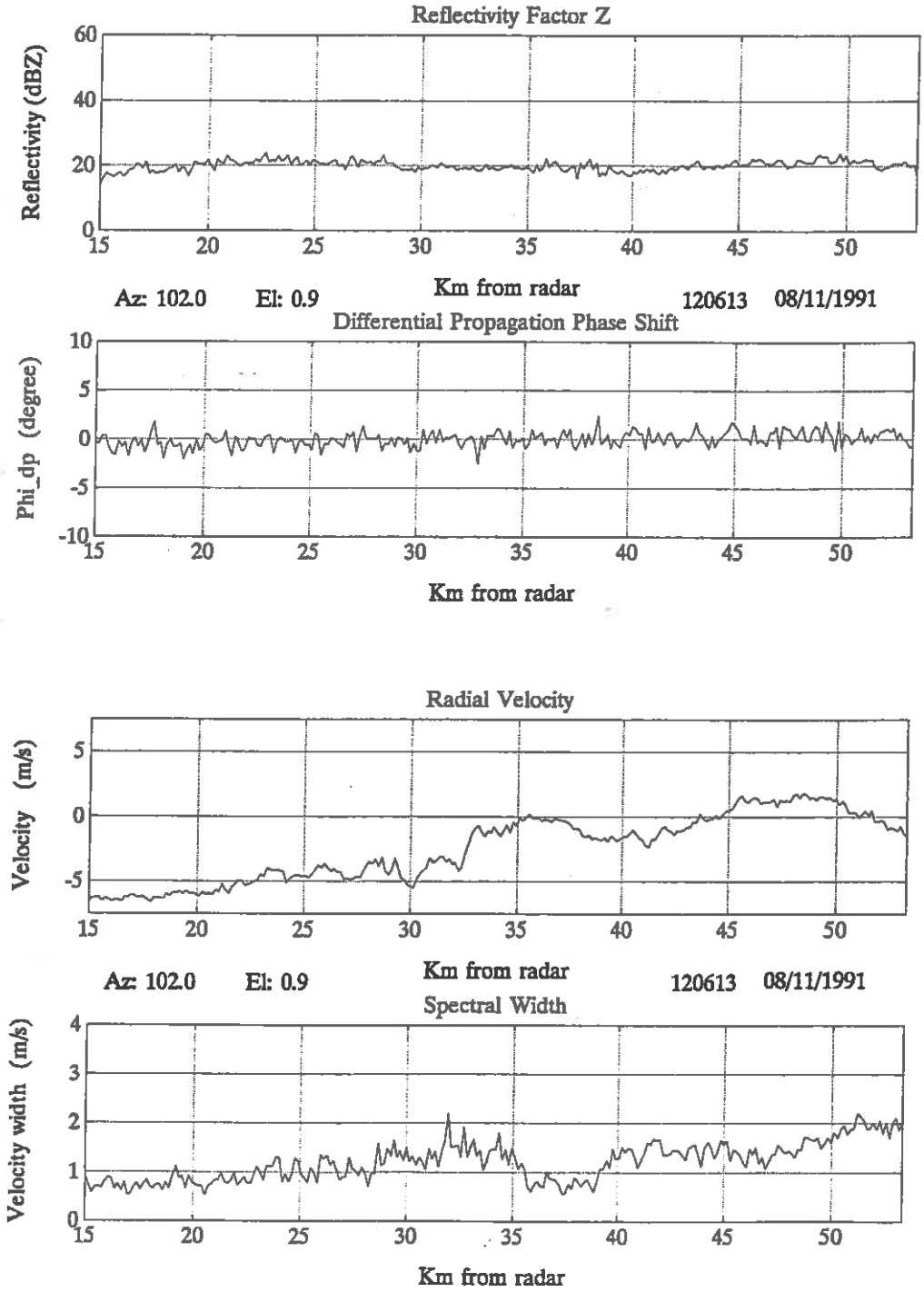


Figure 7 Variation of reflectivity, differential phase, velocity and Doppler width with range for a 0.4 second dwell in light rain.



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